FM Terminal Transmitter and Receiver For the TH Radio System

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The FM terminals form an important subsystem of TH radio, as the link between the 0–10 mc baseband signal and the 74.1-mc FM signal. Severe requirements arise from the design objective of 16 terminal pairs in tandem in 4000 miles. The FM transmitter design uses a 6-kmc reflex klystron as a frequency modulator, the output of which is heterodyned down to 74 mc by another 6-kmc source. Automatic frequency control, with a 74.13-mc crystal oscillator as reference, provides the required frequency stability. In the FM receiver, an IF amplifier-limiter is followed by a balanced FM discriminator which uses parallel resonant discriminator networks. Improved linearity is obtained by special design of a common interstage network. The video amplifiers are balanced and use high-performance electron tubes. The over-all gain of a terminal pair is 8 db, between 124-ohm balanced video circuits.

I. INTRODUCTION

The relation of the FM terminals to the over-all TH system is described briefly in a previous paper.[1] There are two basic types of terminals: an FM transmitting terminal, which converts the baseband signal into a frequency modulated signal centered at 74.1 mc; and an FM receiving terminal, which recovers the baseband signal from the FM signal. The types of baseband signals to be transmitted[1] are shown in Fig. 1. In each terminal appropriate amplification is provided at both the intermediate and baseband frequencies to permit interconnection with other parts of the TH system. FM terminals are required at the ends of a TH route and at intermediate points where the baseband signal, or some portion of it, must be added or dropped. The design is based on a maximum of 16 terminal pairs in tandem in 4000 miles.

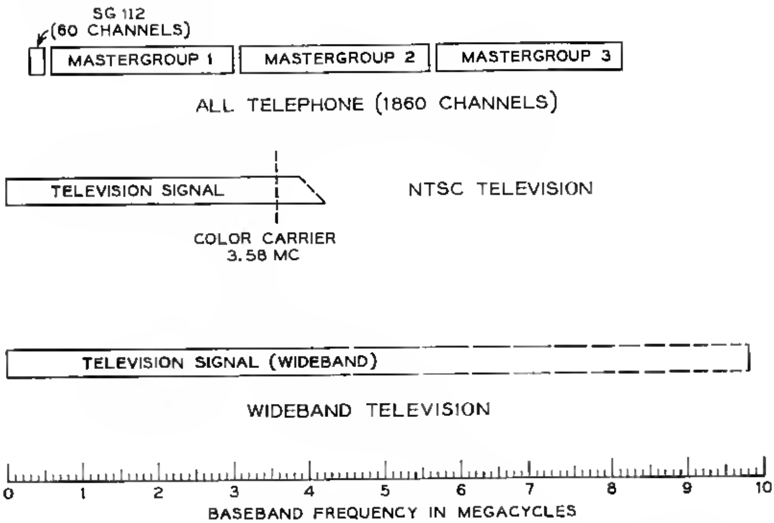

Fig. 1 — The three types of baseband signals transmitted in the TH system.

II. DESIGN OBJECTIVES AND CONSIDERATIONS

2.1 *Over-all Telephone and Television Objectives*

The over-all telephone (TP) and television (TV) objectives for 4000 miles of TH are based on extensive experience with previous systems. From these over-all objectives, allocations were made early in the TH development to the various portions of the system based on estimates of the relative difficulty and expense which would be required to achieve them. The allocations to the FM terminals are summarized in Table I.

Table I — Objectives for 16 Terminal Pairs in Tandem

Telephone (TP)	
Total noise (0 db TL)	31 dba
Fluctuation noise	28 dba
Cross modulation noise	28 dba
Television (TV)	
Weighted signal-to-noise ratio*	62 db
Differential phase	±1.25 degrees
Differential gain	±0.6 db

* The weighted signal-to-noise ratio is defined as the ratio of the peak-to-peak signal to the weighted rms noise, where the weighting is a function of noise frequency. For a detailed discussion see Ref. 2.

2.2 *Signal Characteristics*

Objectives such as those given in Table I can often be expressed in alternative ways. For example, the linearity objective necessary to restrict cross modulation noise is stated above in terms of the resulting noise in dba. For laboratory work, however, it is useful to state the linearity objective in terms of the harmonic performance when a sine wave of a given amplitude is applied. Conversions such as this depend upon certain system parameters. Assumptions, subject to some change when the over-all performance of the system could be determined, had to be made regarding signal characteristics such as the peak frequency deviation and the amount of pre-emphasis. The values of these quantities are shown in Table II.

2.3 *Objectives for an FM Terminal Pair*

Objectives for a single terminal pair were obtained by reducing the total allocation of Table I by a factor of either 4 (12 db) or 16 (24 db); the choice depends on whether a particular impairment could be expected to add on a random or on a systematic basis.

The principal design objectives for a single terminal pair are shown in Table III. In the sections which follow some of them will be discussed briefly.

2.4 *Baseband Transmission*

Essentially flat transmission is desired over the frequency band of all three types of signal. The upper frequency limit is approximately 10 mc for two; the lower frequency limit is set by the 60-cps component of a television signal. To provide adequate phase linearity for the low-frequency TV components, it is necessary to keep all low-frequency cutoffs considerably below 60 cps. Based on experience with other video systems, a low-frequency objective is expressed in terms of the distortion to a 60-cps square wave. A distortion not exceeding 2 per cent of the nominally flat top of the square wave is considered acceptable for a terminal

TABLE II — SIGNAL CHARACTERISTICS

Telephone signal	
Peak frequency deviation	4 mc
rms frequency deviation	0.7 mc
Pre-emphasis	7.5 db
Television only	
Peak frequency deviation	4 mc
Pre-emphasis (tentative)	12 db

TABLE III — DESIGN OBJECTIVES FOR ONE FM TERMINAL PAIR

Baseband transmission	
Bandwidth	2 cps to 10 mc
Gain stability	±0.25 db
Peak frequency deviation	4 mc
Center frequency of FM transmitter	74.13 ± 0.1 mc
Harmonic performance*	
Peak deviation for applied sine wave	4 mc
Second harmonic with respect to fundamental	−49 db
Third harmonic with respect to fundamental	−51 db
Differential gain*	See Note
Differential phase	0.3 degree
Fluctuation noise	
Message (at 0 db TL)	16 dba
Television (weighted signal-to-noise ratio)	74 db

* Differential gain and harmonic performance are analytically related as shown in Appendix A. The design objectives above for harmonic performance are set by cross modulation requirements for telephone; they also insure adequate differential gain performance for television.

pair. To meet this, the low-frequency cutoff (3-db point) actually occurs at about 2 cps. Transmission flatness of about ±0.1 db is the objective for the band from 60 cps to 10 mc.

The objective of ±0.25 db for gain stability comes from two main considerations. First, there is the desire to control net loss in toll telephone channels within rather close limits. Second, all operating terminals at a given point must have almost the same net loss as standby terminals. Otherwise, switching from a regular terminal to a protection terminal will cause hits in data signals.

2.5 *Harmonic Performance*

Nonlinearity in the terminals causes cross modulation in the TP signal, and differential phase and gain in the TV signal. It was found that the linearity required for the telephone signal was controlling by a slight margin. This led to the objective on harmonic performance given in Table III.

2.6 *Differential Gain and Phase*

The NTSC color television signal uses a modulated 3.58 mc carrier to transmit color information. Intermodulation between the low-frequency luminance information in the signal and the color carrier causes amplitude and phase variations in color carrier which are a function of the luminance signal. These variations show up as distortion in the saturation and chroma of the reproduced TV picture.

A special test signal[3] consisting of a low-level 3.58-mc tone and a higher-level 15.75-kc tone is used to simulate the TV signal and to test for distortion of this type. Variations in the amplitude and phase of the 3.58-mc tone as a function of the 15.75-kc tone are referred to as differential gain and phase. For tests of this type the peak-to-peak amplitude of the low-frequency tone is normally made equal to the peak-to-peak amplitude of the TV signal it simulates. It is then possible to specify quantitatively the amount of differential gain and phase which corresponds to a tolerable amount of color distortion in the TV picture.

Since differential gain and phase distortion occurs as the result of intermodulation, the amount of distortion which occurs in a particular nonlinear system depends on the amplitude of the applied signal. Differential phase and gain distortion can be reduced by reducing the signal amplitude, but at the expense of a poorer signal-to-noise ratio. A compromise between these two types of signal degradation is often possible by reducing the amplitude of the low-frequency components of the TV signal before transmission by means of a pre-emphasis network. An inverse network is used to compensate at the receiving end. Pre-emphasis of the amount shown in Table II is proposed for this system. With this pre-emphasis, the objectives for harmonic performance shown in Table III ensure that the over-all differential gain objectives will be met.

III. FM TRANSMITTER

The FM transmitter provides a +11-dbm output signal, centered at 74.1 mc, which is frequency modulated proportional to the input baseband signal. One volt peak-to-peak signal produces 8-mc peak-to-peak frequency deviation.

To preclude excessive intermodulation, very little nonlinearity is permitted in the frequency deviation vs voltage characteristic. This objective had a very strong influence on the selection of a modulation method. Other objectives that had important influences on detailed design approaches were: the wide baseband, a conversion gain stability of ±0.15 db, and a carrier frequency (74.13 mc) stability of ±0.1 mc.

After a considerable exploratory development period during which several more compact circuits were rejected because of marginal linearity or bandwidth, a reflex klystron was selected as the frequency modulator. A similar circuit using klystrons is successfully employed in the FM modulator associated with the TD-2 radio system.[4] However, improvements in the klystron and its associated circuits were essential to meet the more stringent requirements of the TH system.

3.1 *General Description*

Selection of a klystron modulator essentially establishes the principal accessory units. These are shown on Fig. 2.

The baseband signal, amplified by the video amplifier and applied to the repeller of the deviated (DO) klystron, causes its frequency to vary around a rest value of approximately 6174 mc. The deviated signal is applied through an isolator to the converter, where it is mixed with a 6100-mc signal from the beating (BO) klystron. The output, a frequency modulated wave centered on 74.1 mc, passes through a delay equalizer (which compensates for the delay distortion of a tandemly connected FM transmitter and FM receiver), to an IF amplifier. The amplifier output connects via coaxial cable to the other parts of the TH system.

To provide an input for the automatic frequency control (AFC) circuit, a small fraction of the output from the FM transmitter is abstracted and amplified. Alternating samples of this FM signal and the output of a crystal-controlled 74.1-mc oscillator are applied to the limiter-discriminator. The output is a square wave with an amplitude proportional to the frequency difference. This error signal is amplified and rectified in the synchronous detector. The resultant voltage is applied to the BO klystron with the proper polarity to reduce the average frequency error.

A photograph of the FM transmitter is shown in Fig. 3. The klystrons, converter and other microwave devices are mounted in the FM generator panel. The video amplifier is at the bottom and the transmitting IF amplifier is just above the frequency comparator panel. A further description of the equipment features is given in a companion paper.[5]

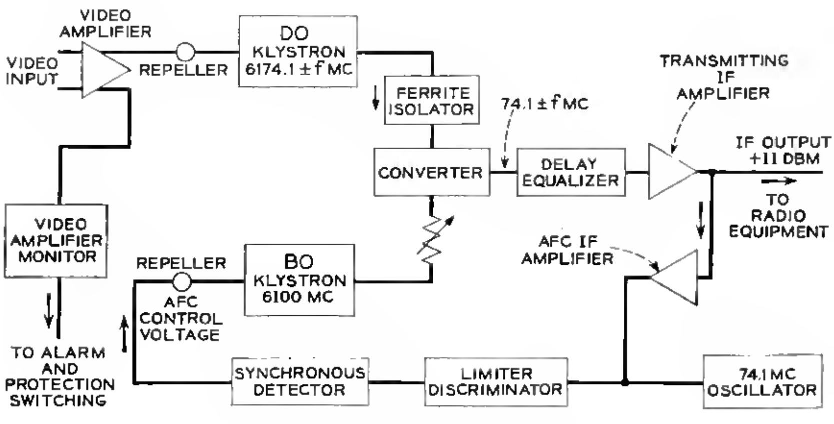

Fig. 2 — Block diagram of FM transmitter.

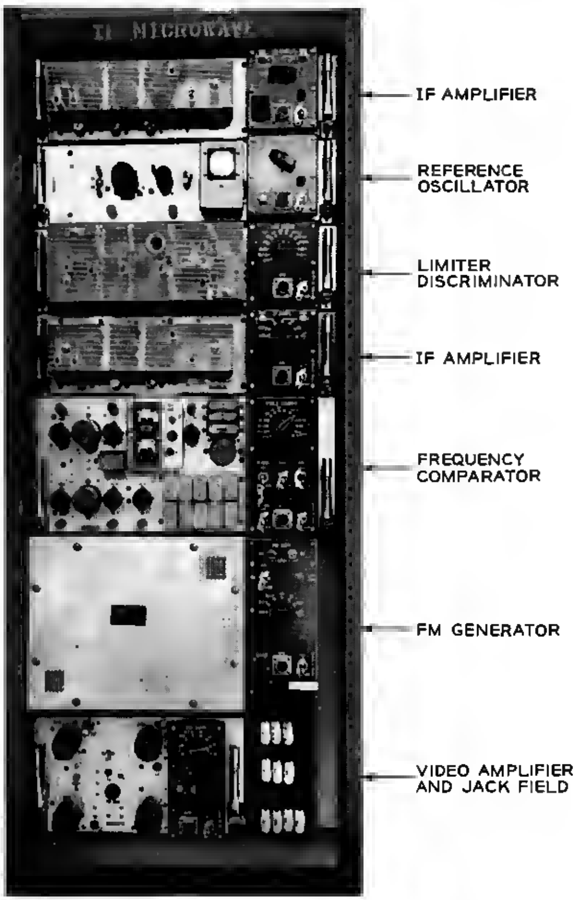

Fig. 3 — The FM terminal transmitter.

3.2 *Video Amplifier**

The gain of the video amplifier is such that a signal of 1 volt peak-to-peak applied to its 124-ohm balanced input is increased to 5.5 volts peak-to-peak at the repeller electrode of the DO klystron, which is sufficient to deviate the klystron 8 mc peak-to-peak. At this output voltage, a typical video amplifier has second and third harmonics that are below the fundamental by 60 db and 75 db respectively.

To realize this performance, two balanced electron-tube amplifier

* Important contributions were made to this and other sections dealing with the video amplifiers by H. C. Hey.

stages are used in a circuit configuration similar to that of the video amplifier described in Section IV. Low second- and third-harmonic distortion results from the use of relatively lightly driven, high-current electron tubes in the two stages. Western Electric 448A tetrodes and 437A triodes are used, respectively, in the input and output stages. Additional suppression of the second harmonic is achieved through the balancing action of push-pull stages, which have cathode feedback to reduce and stabilize residual unbalances.

To keep the gain-frequency distortion less than ±0.1 db up to 10 mc, design emphasis is placed on minimizing spurious interstage capacitance. Interstage resistances are limited to values which give a 3-db gain reduction at 14 mc. Interstage compensation, part of which is individually adjusted, is used to achieve flat gain. The use of large coupling capacitors, along with some additional phase compensation, limits 60-cps square wave distortion to less than 1 per cent.

3.3 *Microwave Circuit*

The microwave circuit consists of the two klystrons, an isolator, an RF attenuator and a converter. The converter employs a silicon diode in an unbalanced microwave network with internal resistance padding to reduce changes in performance with different diode characteristics. The output power from the klystron is sufficiently high to make unimportant the added conversion loss due to the padding. Variations in the output amplitude of the FM wave are reduced by the limiting action obtained in the converter. This limiting action is achieved by making the power from the frequency-deviated oscillator (DO) substantially higher than that from the undeviated beating oscillator (BO). The RF attenuator determines the level difference, since the two klystrons are of the same type and generate approximately the same power output. Presentation of a well-matched impedance to the DO klystron is required for modulation linearity, and is obtained by use of the microwave ferrite isolator, as shown in Fig. 2.

Radiated microwave interference to or from the klystrons is precluded by enclosing them in a shielded compartment into which all leads are brought through microwave filters.

3.4 *Reflex Klystron Modulator*

To meet the stringent linearity objectives for the TH system a new klystron, coded as the Western Electric 450A, was designed. It has a limited tuning range (6000 mc to 6200 mc), a low-Q resonator (loaded

$Q \approx 100$) and requires resonator, repeller and heater potentials of approximately 450, -100 and 6.3 volts, respectively.

The loaded Q of the klystron is important in determining both the linearity and the fluctuation noise performance. As shown in Appendix B, the principal distortion terms due to nonlinearity in the voltage-frequency deviation characteristic are proportional to Q^2. On the other hand, fluctuation noise, due to shot noise in the electron stream, is proportional to $Q^{-1/2}$ (Appendix D). Thus, the selection of too high a Q leads to excessive intermodulation products, whereas too low a Q causes excessive fluctuation noise. A compromise value of 100 was selected to give the best over-all performance. With this, the second and third harmonics are respectively at least 54 db and 56 db below a fundamental which has a peak-to-peak deviation of 8 mc. The FM components of the fluctuation noise, as measured by an FM receiver, are flat with frequency above about 25 kc. Below 25 kc the noise power increases with an approximate $1/f$ law. Above 25 kc the rms frequency deviation due to the noise in two klystrons is approximately 0.6 cps in a one-cycle band.

The deviation sensitivity of the klystron is approximately inversely proportional to loaded Q, and essentially independent of baseband frequency up to 10 mc. The reasons for this are demonstrated in Appendix C.

3.5 *IF Amplifier*

The two identical IF amplifiers (transmitting and AFC) shown in Fig. 2 each have three tubes. Each has a minimum gain of 21 db, a maximum output of $+11$ dbm, and a bandwidth of 58 to 90 mc between 0.3-db points. The bias on the intermediate tube, supplied from an external source, can be varied for manual gain adjustment or for gating the amplifier on and off as required in the AFC circuit.

Input, output and intermediate stages in this amplifier are almost identical in electrical design to corresponding stages in the main IF amplifier of the radio receiver, described in a companion paper.[6]

3.6 *AFC Circuit*

The AFC circuit, shown schematically on Fig. 4, is designed to hold the average frequency of the outgoing FM wave at 74.1 ± 0.1 mc. A low-temperature-coefficient, crystal-controlled oscillator operating at 74.130 mc provides the basic reference against which the average frequency of the outgoing FM wave is compared. By alternately gating the 74.1-mc reference oscillator and the AFC IF amplifier on and off at a

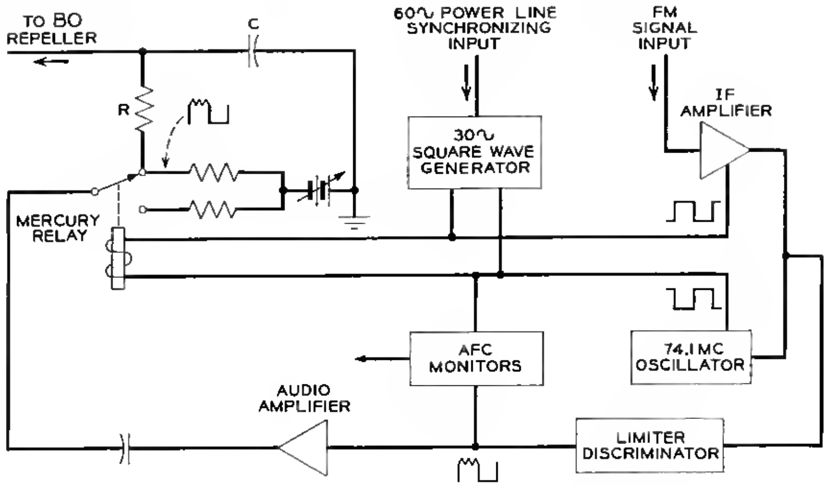

Fig. 4 — Functional schematic of the AFC system.

30-cps rate, there appears at the IF limiter input during one $\frac{1}{60}$-second interval a signal from the 74.1-mc oscillator, and during the next $\frac{1}{60}$-second interval a sample of the outgoing FM signal. The limiter section equalizes the signal amplitudes, and the frequency discriminator delivers to the audio amplifier a 30-cps square wave whose peak-to-peak average amplitude is proportional to the frequency difference between the reference oscillator and the average (carrier) frequency of the outgoing FM wave. After amplification the 30-cps wave is rectified in a synchronous detector (the mercury relay), the dc output of which is related in amplitude and polarity to the frequency difference. After filtering, this dc is applied to the repeller of the BO klystron with a polarity which causes its frequency to change in a direction to reduce to a small value the difference (error) in frequencies of the 74.1 mc oscillator and the outgoing FM wave.

Since the instantaneous frequency of the FM wave is continuously changing, nonlinearity in the AFC discriminator will give an average output not linearly related to the average frequency of the FM wave. This discriminator has a first-order nonlinearity of 2 per cent or less between 70 and 78 mc. On the basis that the rms frequency deviation will be less than ± 1 mc, the average frequency shift error will be less than 5 kc.

The use of ac amplification following the discriminator makes the circuit insensitive to small drifts in the center frequency (corresponding to

a dc output of zero) of the discriminator. Synchronous rectification of the amplified ac output restores sense so that the resultant voltage is applied to the BO repeller with the correct polarity. Since the AFC open-loop gain is typically 40 db, uncorrected frequency differences of as much as 10 mc are reduced by AFC action to 100 kc.

When the outgoing FM wave contains a television signal, large 60-cps and harmonically related voltages are present in the discriminator output during signal sampling intervals. A dc output related to the true average values of the signal voltage must be obtained for accurate frequency control. This is accomplished by using negative feedback to linearize the audio amplifier and by using a mercury relay switch for rectification. The switch has negligible storage reactance in its output load so that its dc output is proportional to the average value of the input.

When the frequency of the local TH ac power differs significantly from that of the remote power line against which the television signal is synchronized, successive signal samples begin (and end) at different phases of the 60-cycle ac component in the TV signal. Consequent "flicker" interference[7] on television signals caused by the apparent shifts in average frequency at the beat frequency rate are reduced to tolerable values by the RC filter following the rectifier.

Other beat frequency effects are kept small by synchronizing the 30-cps gating and rectification functions with local power frequencies. The RC filter on the rectifier output has a 3-db cutoff at 0.005 cps. It gives 60-db loss at 5 cps, the most annoying beat frequency, and reduces the 40-db AFC loop gain to unity at 0.5 cps. Transient response is optimized by virtue of the 90° phase asymptote. Video phase shift, and consequently the 60-cps square wave response of the FM transmitter, is only slightly affected by the AFC action, which has an effect approximately equal to that of a low-frequency cutoff at 0.5 cps.

IV. FM RECEIVER

The FM receiver accepts the 74-mc FM signal and delivers a balanced baseband output signal which is 8 db above 1 volt peak-to-peak in a 124-ohm circuit for 8-mc peak-to-peak frequency deviation.

Performance requirements on linearity, baseband transmission, stability and noise are comparable to those already discussed for the FM transmitter. After careful study, the design described in the following sections was selected as the best compromise between over-all performance and ease of maintenance and adjustment. In its main features the design is similar to that which has been used in the TD-2 radio system[4] for a number of years. Modifications in the detailed circuitry, however, have

achieved a considerable improvement in bandwidth, linearity and circuit stability.

4.1 *General Description*

A block diagram of the FM receiver is shown in Fig. 5. The FM input is applied to the receiver by means of 75-ohm coaxial cable. Provision is made for an IF amplifier, identical with the two in the FM transmitter, in case additional IF gain is required. In any event, the FM signal at +1 dbm is applied to an amplifier-limiter to suppress any amplitude modulation of the signal, which otherwise would cause unwanted distortion in the discriminator. In addition, the limiter action tends to maintain a constant input power for the discriminator circuit. Without it, the discriminator output would vary linearly with changes in input carrier power to the FM receiver. This would cause undesirable variations in the net gain of a terminal pair. The amplifier-limiter is electrically identical to the one in the broadband radio transmitter.[6] The mechanical design, however, is somewhat different,[5] to be in keeping with the plug-in styling of the other FM terminal circuits.

The adjustable attenuator ahead of the discriminator is used to set the sensitivity of the discriminator section to a standard value. The discriminator recovers the baseband signal from the FM wave, in two steps. First, networks which introduce amplitude slope across the IF band produce amplitude modulation, which is proportional to the frequency modulation of the input signal. Amplitude detectors then recover the baseband signal, which is subsequently amplified in the baseband amplifier.

A photograph of an FM receiver is shown in Fig. 6, in which the individual units are easily identified.

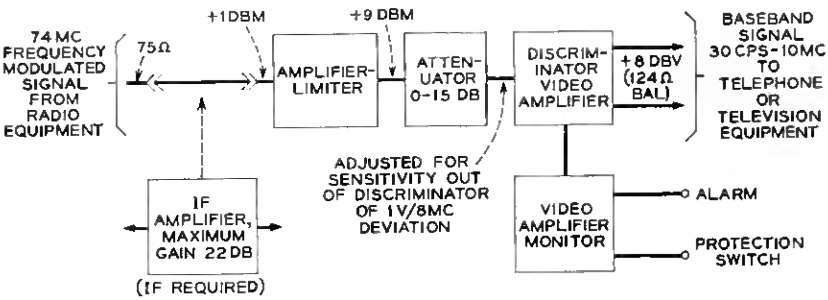

Fig. 5 — Block diagram of FM receiver.

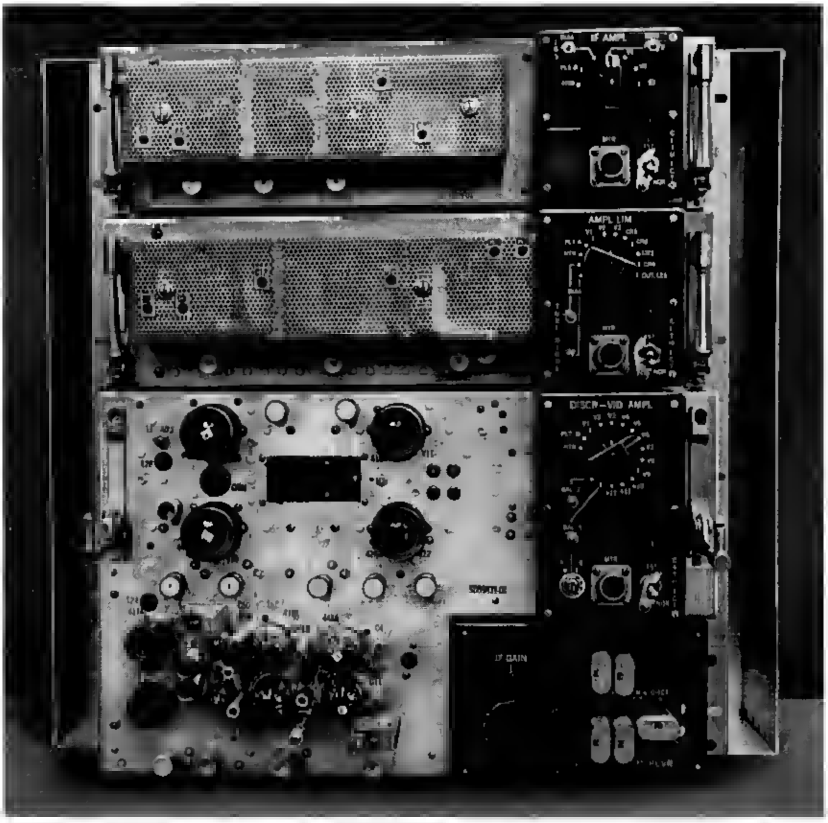

Fig. 6 — The FM terminal receiver.

4.2 *FM Discriminator and Video Amplifier*

Simplified schematics of the FM discriminator and of the video amplifier are shown in Figs. 7 and 8. The input signal is transformer-coupled to two 448A electron tubes, v_1 and v_2, which are operated in parallel. Their combined output is developed across a common interstage, consisting essentially of a parallel resonant circuit, and applied to the grids of the 418A electron tubes, v_3 and v_4, which are driven in parallel. All four tubes are stabilized by dc feedback and cathode compensation networks (z_1—z_4) as described for the IF amplifier in the broadband radio receiver.[6] The common interstage performs an important function in the over-all design which will be described later. A 75-ohm test jack is provided to aid in interstage adjustment.

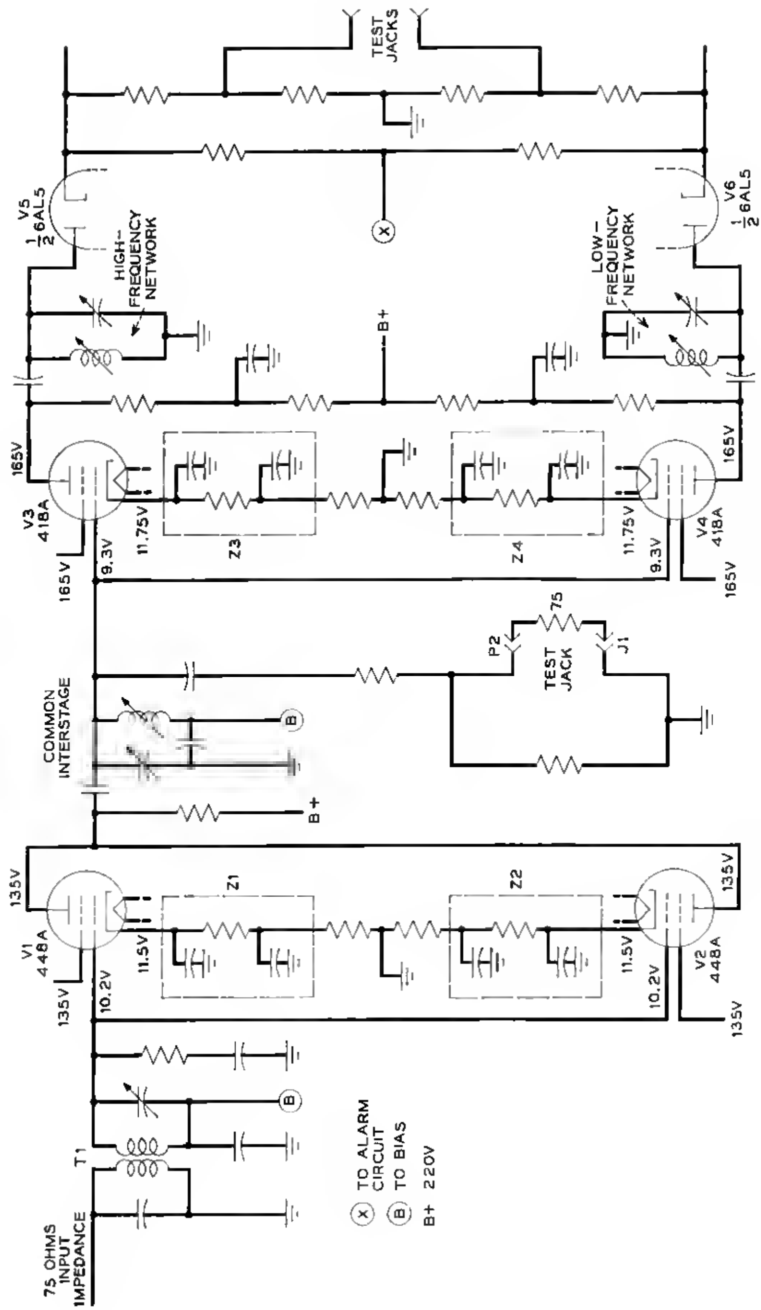

Fig. 7 — Simplified schematic of FM discriminator.

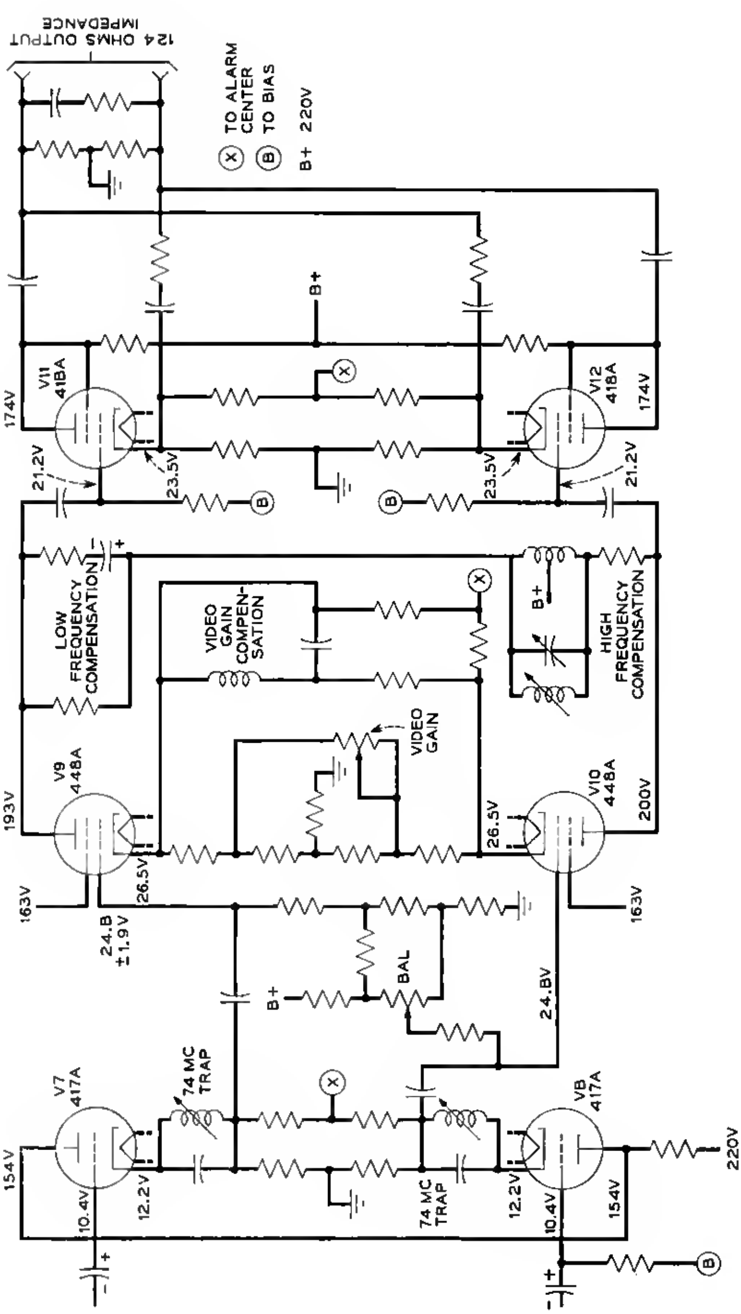

Fig. 8 — Simplified schematic of video amplifier.

Tubes V3 and V4 are used as constant current generators to drive separate parallel-resonant discriminator networks. The high-frequency network is peaked at approximately 100 mc, and the low-frequency network is peaked at approximately 50 mc. Each provides a large amplitude slope across a wide band centered at 74.13 mc. Tuned at these frequencies, the slope of one is positive and the other negative across the IF band. The output voltages developed across the networks are amplitude modulated with 180° phase difference between them. The diode detectors, V5 and V6, recover the baseband signal from the amplitude modulation, thus providing a balanced input to the balanced video amplifier which follows.

The video amplifier section (Fig. 8) consists of three balanced stages in tandem. The first is a cathode follower stage employing 4t7A triodes. The cathode follower is used to minimize the capacitance facing the diode detectors. Parallel resonant traps are provided in the cathode circuits to reduce the amount of 74-mc carrier entering the amplifier stages which follow. The second stage, with 448A tetrodes, provides the required voltage amplification and feeds the balanced output stage. Fixed low-frequency compensation and adjustable high-frequency compensation are provided in the interstage. The output stage, using 418A tetrodes, is connected as a modified cathode follower circuit. The output is increased by connecting both cathodes and plates to the load in an arrangement which is used in the A2A television transmission system.[8]

4.3 *Design Considerations*

To meet the over-all system objectives, the discriminator has to be very linear, yet give an output large enough to provide an adequate signal-to-noise ratio at the input to the video amplifier. Furthermore, the circuit must be readily adjustable to allow for manufacturing variations and subsequently be stable with time. The manner in which the design has been affected by these considerations is discussed in the following paragraphs, in which the usual order is reversed by working from the output toward the input.

A study of the noise and microphonics expected in the first stages of the video amplifier leads to the establishment of an objective of 1 volt peak-to-peak at the detector output, for a peak-to-peak frequency deviation of 8 mc. This level is sufficient to keep the fluctuation noise contribution of the FM receiver at least 10 db below that of the FM transmitter and at the same time to prevent microphonics in the first stages of the video amplifier from degrading the television signal.

The use of 6AL5 diodes for the AM detectors provides a compromise

among the following objectives: good linearity, high detection efficiency and ease of replacement. The operation is between that of an averaging detector and a peak detector. The over-all detection efficiency is approximately 40 per cent, close to that of an ideal averaging detector. Thus, the capacitance in the output circuit for the diodes provides some peaking action to compensate for the loss due to the forward resistance of the diode. However, this capacitance must be kept low to minimize video roll-off at 10 mc.

From the desired output of 1 volt peak-to-peak and the diode efficiency of 40 per cent, the necessary change in IF signal amplitude as it is tuned across an 8-mc band is about 1.25 volts for each discriminator network. This change in signal amplitude is a function of the discriminator networks and the signal currents provided by the preceding tubes. Restrictions are imposed by the linearity objective and the interstage capacitance which must be absorbed. This limits the maximum change in impedance which can be achieved across the 8-mc band to about 50 ohms. Thus, peak signal currents of about 25 ma are required from the driving tubes. Furthermore, this amplitude must be provided with low harmonic content. The second harmonic, in particular, will be enhanced with respect to the fundamental by the amplitude-frequency characteristic of the high-frequency network. The detector output will therefore contain an error term due to the harmonics. Good harmonic performance is required to permit accurate adjustment of the discriminator with sweep signals, and to a somewhat lesser extent, to prevent distortion to the normal signal.

The need for a large signal current with low harmonic content led to the selection of the 418A tube for this application.

4.4 *Discriminator Linearity*

The discriminator networks have a substantial amount of curvature, predominantly parabolic, as shown in Fig. 9. This curvature, if uncompensated, would result in a nonlinear relationship between the incoming frequency modulation and the resulting amplitude modulation. One method of correction is to use more complex discriminator networks. This was not selected because of the difficulty in controlling parasitic capacitance and inductance. A second approach, used in the discriminator for the TD-2 system, is to select network designs such that the parabolic curvatures of the two sides are equal. The predominant second-order modulation products then tend to cancel each other in the balanced output from the detectors. The major difficulty with this approach is the amount of balance required. An analysis of the networks in the

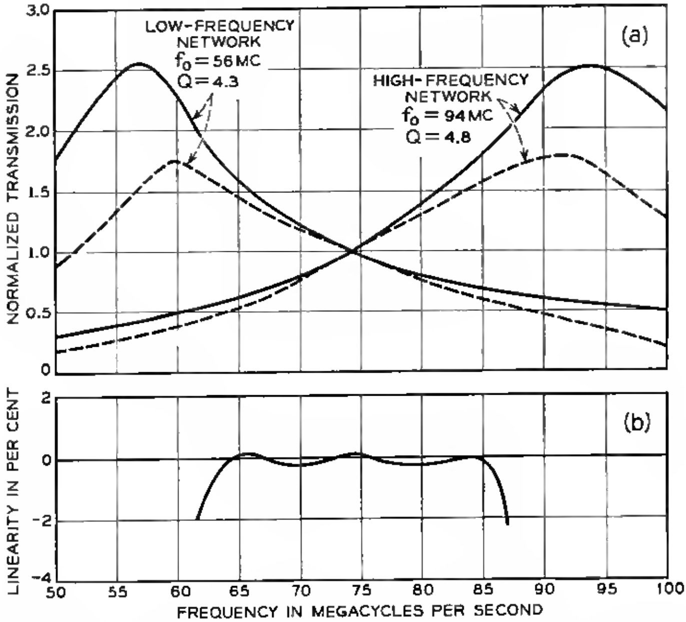

Fig. 9 — (a) Discriminator network characteristics; dashed curves show effect of common interstage; (b) typical discriminator linearity.

TH discriminator indicated that the second-order modulation products from one side of the discriminator would be about 30 db above the desired objective. Although a 30-db balance is obtainable at the time of adjustment, it is difficult to hold between maintenance intervals since it depends on the relative gains of V3 and V4.

Another approach,* the one used in this design, also requires the networks to have equal parabolic curvatures. However, instead of depending on precise cancellation in the output, a compensating parabolic shape of opposite sign is introduced in the common interstage ahead of V3 and V4. In this way the linearity of each side of the discriminator is substantially improved, as shown in Fig. 9. With this design the nonlinearity of each side exceeds the over-all objective by only about 10 db. This reduces by a considerable amount the effect of gain changes in the driving stages V3 and V4. For example, without the common interstage, only 0.5 db of change in the relative gains after the initial balance would

* This approach was suggested by N. E. Chasek of the Radio Research Department at Bell Telephone Laboratories.

cause the objective to be exceeded. With the common interstage, the acceptable variation is 4 db. Thus, the common interstage substantially increases the time stability of the second harmonic balance.

The procedure used in designing the discriminator networks is outlined in Appendix E. This design was modified slightly on the basis of experimental results to obtain the typical linearity characteristic shown in Fig. 9(b).

V. POWER SUPPLIES

Supplied with regulated, reliable 220-volt, 60-cycle ac inputs, the power rectifiers give dc voltages of -11, $+135$ and $+220$ volts for the FM receiver; -11, -170, $+135$, $+220$ and $+450$ volts for the FM transmitter. Electronic regulation is counted on to reduce bobble and consequent television flicker interference to acceptable levels. This stability also keeps within limits, for extended time intervals, variations in FM deviation, FM sensitivity, and second harmonic balances. Additional details of the power supplies are given in Ref. 9.

VI. MONITORS

To preclude disruption or excessive degradation in service due to failure in an FM transmiter or an FM receiver, the failed unit is automatically replaced by a standby unit. Status information for initiating this protection switching action and for registering alarms is obtained from monitors on space current in video amplifier tubes, on IF carrier power at the FM transmitter output, on three significant parameters in the AFC system, and on rectified carrier level in the FM receiver.

6.1 *Video Amplifier Monitors*

A considerable simplification in instrumentation of video amplifier monitors is based on the fact that any reduction in gain, or increase in harmonic distortion, will probably be accompanied by a change in space current in one or more electron tubes. Cathode voltages (which are proportional to space currents) from all tubes are added, and the sum is applied to the input of a differential dc amplifier. A change in this sum resulting from a change in any cathode voltage of 30 per cent or greater will initiate an alarm and a protection switching order. Even though the response time of this monitor is around two milliseconds, it is not operated by television waveforms having frequency components as low as 60 cps because of the push-pull action in the balanced video amplifier.

6.2 *IF Carrier Monitors*

Failures in klystrons or the transmitting IF amplifier in the FM transmitter are detected by a reduction in output from an IF carrier level detector connected across the outgoing IF line.

By monitoring the sum of rectified voltages at the discriminator output in the FM receiver, information is obtained on the status of IF carrier input and of the electron tubes and circuits in the discriminator. This voltage sum is combined with the sum of the FM receiver video amplifier cathode voltages and applied to a video amplifier monitor circuit of identical design to that used in the FM transmitter.

6.3 *AFC Monitors*

Because of the large time constant in the output filter, a failure in the AFC control circuit will not cause an immediate change in the average frequency of the outgoing FM wave. This allows the use of three reliable, though relatively slow-acting, sensitive meter relays for monitoring this circuit. These relays monitor: (a) the peak-to-peak frequency error voltage, (b) the rectified carrier level, and (c) the 30-cps gating voltage. Failures in the audio amplifier or synchronous rectifier will cause monitor (a) to initiate alarm and protection switching orders whenever the klystrons drift sufficiently to create a 1.5-mc shift in difference frequency. However, monitor (a) is a null indication and so will not be activated by failures in the gating circuits or in the limiter-discriminator and preceding IF circuits. Failures in these circuits are detected by monitors (b) and (c).

VII. TANDEM PERFORMANCE

The baseband-to-baseband performance of a typical FM transmitter and FM receiver when connected in tandem is discussed in this section. This performance, compared with values in Table III, will show how well the design objectives have been met.

It should be noted that "typical" back-to-back performance means the average performance that can be expected during the time intervals between maintenance adjustments on individual FM transmitters and receivers. These specifications draw from experimental data in attempting to give the most probable performance that can be expected over a reasonable time interval after maintenace adjustments have been made. They observe the practical restriction that terminals cannot be connected back-to-back in the field to adjust paired terminals for optimum performance.

7.1 *Envelope Delay Distortion (EDD)*

Measured as the change in phase shift of a 278-kc tone while the IF center frequency is swept between 64 mc and 84 mc, the IF delay distortion (excluding the delay equalizer shown in Fig. 2) has the typical characteristic shown on Fig. 10(a). Odd-order EDD at 64 and 84 mc is approximately -5.3 and $+5.3$ mμs respectively, and even-order EDD is approximately $+8.7$ mμs at both frequencies. To this total the klystron and its microwave circuits contribute almost nothing, the FM discriminator contributes about half, and the remaining IF circuits (amplifier-limiter and transmitting IF amplifier) contribute about half. With the delay equalizer, the distortion is reduced by a factor of at least five.

7.2 *Harmonic Distortion*

Typical curves of second and third harmonic performance (exclusive of the delay equalizer) are shown in Fig. 10(b). At low frequencies the harmonic performance is flat with frequency, and is primarily due to nonlinear voltage-frequency characteristics in the klystron and in the discriminator. At higher frequencies, however, second and third harmonics tend to increase in proportion to the frequencies where the harmonics fall. This is primarily due to EDD in the IF circuits. This source is reduced by delay equalization until it is negligible compared to the low-frequency asymptote. The performance then becomes that shown by the dash lines in Fig. 10(b).

The limiting low-frequency value of -70 db for the third harmonic-to-fundamental ratio comes from systematic contributions by two video amplifiers, the DO klystron and the FM discriminator; typical values are respectively -0.0004 (-68 db), $+0.0015$ (-56 db) and -0.001 (-60 db). The second harmonic contribution from the same four units depends upon the status of balance in each unit. The low-frequency asymptote of -50 db was obtained by taking the root-sum-squares (R.S.S.) of limiting design values of ± 0.001 (-60 db), ± 0.002 (-54 db) and ± 0.002 (-54 db) in each of two video amplifiers, the DO klystron and the FM discriminator, respectively.

7.3 *Differential Gain and Phase*

As discussed previously, the harmonic performance given above insures adequate differential gain performance for television.

Without delay equalization, the differential phase characteristic has the same broad structure shape as the EDD characteristic of Fig. 10(a),

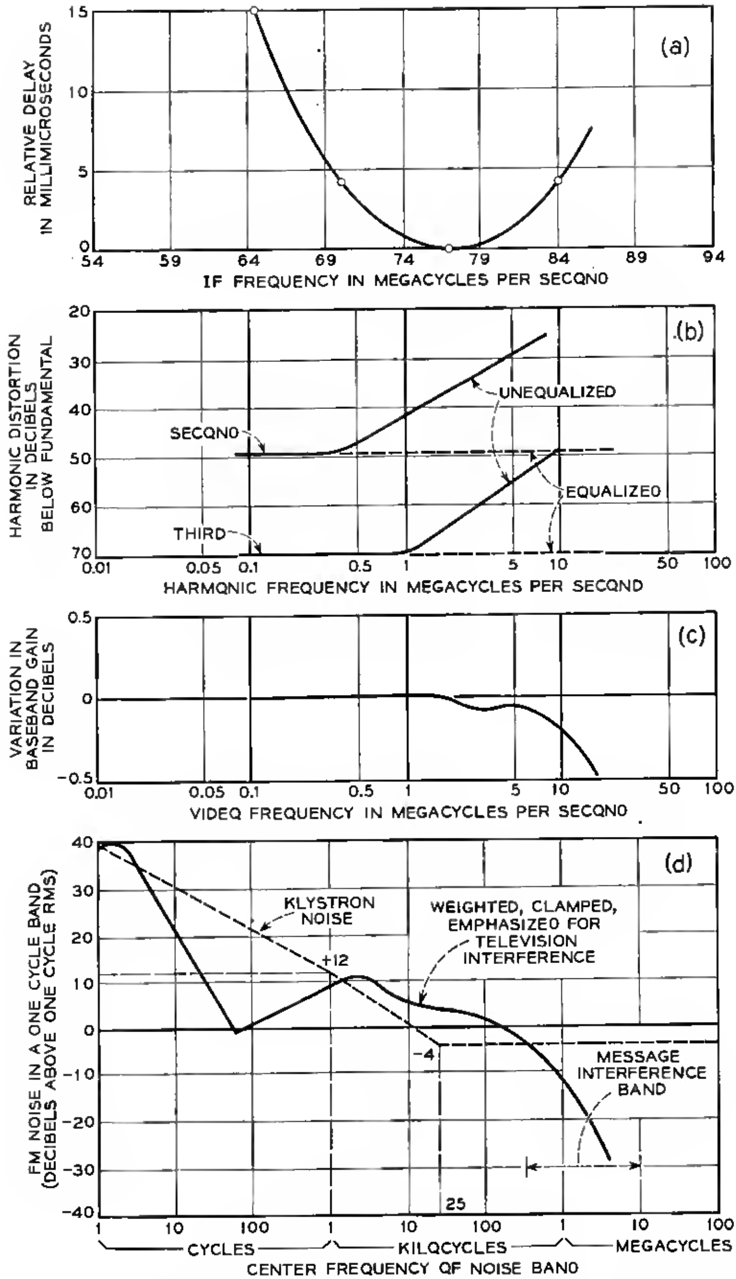

Fig. 10 — Tandem characteristics for an unequalized pair: (a) IF delay distortion, (b) typical harmonic distortion for 8-mc peak-to-peak deviation, (c) typical baseband response, (d) fluctuation noise.

but with reversed sign, since it is also the change in phase shift (in this case at 3.58 mc) while the carrier is swept by a low-frequency (15-kc) video voltage. Odd-order and even-order differential phase are each about 2.5° at ±4-mc peak deviation. Since the low-frequency components of the TV signal are less than this by 12 db due to pre-emphasis, the odd- and even-order differential phase experienced by the TV signal is reduced by factors of at least 4 and 16 respectively. Differential phase is further reduced by the delay equalization, and hence is well below the 0.3° objective.

7.4 *Baseband Amplitude Response*

Low-frequency baseband response, determined by the characteristics of the two video amplifiers, is typically down 3 db at 2 cps. This, with some added low-frequency phase equalization, gives a 60-cps square wave response having low-frequency slope of less than 2 per cent. High-frequency baseband response, typically ±0.15 db at 10 mc, is the result of a number of significant contributions.

As described in previous sections, video response in the klystron and its microwave circuits is substantially constant. Response in each video amplifier is adjusted to be constant up to 10 mc within ±0.1 db.

The EDD of Fig. 10(a) produces a video roll-off of 0.06 db at 10 mc. In contrast to harmonic distortion, amplitude response is made worse by delay equalization when the distortion and the equalization are separated by a limiter. Thus, equalization in the FM transmitter, which introduces an equal and opposite delay distortion, adds another 0.06 db to the video roll-off at 10 mc.

Contributing directly to baseband gain-frequency variations, even-order IF gain variations in a typical transmitting IF amplifier account for ±0.03 db. Though somewhat more subtle to observe experimentally because of limiting action, the same basic mechanism in the amplifier-limiter contributes approximately the same video gain deviation. Linearity and detector response in the FM receiver results in a systematic even-order characteristic of about −3 per cent at 64 mc and 84 mc, contributing −0.3 db to the video response at 10 mc.

Summing up the systematic contributions and adding the random contributions on an rss basis leads to a typical characteristic of −0.4 ± 0.15 db at 10 mc. The systematic characteristic is equalized to give the nearly flat response shown on Fig. 10(c).

7.5 *Low-Frequency Noise*

Noise below 300 kc causing interference in TV signals comes from mechanical vibration in electron tubes, from "$1/f$" cathode emission

fluctuation noise in klystrons, from power rectifier output voltages having "hum" components which are multiples of power-line frequency, and from "bobble" components which are random variations in power rectifier dc outputs at rates of 1 to 30 cps.

Rugged mechanical design in the klystrons and maintenance of relatively high signal levels at the FM receiver video amplifier input minimize mechanical vibration noise. Considering frequency weighting, TV circuit clamper characteristics, and 12-db pre-emphasis for tbe combined service, the contribution by "$1/f$" cathode emission noise over the band of 0 to 25 kc can be neglected since it is more than 6 db below weighted rms fluctuation noise in the rest of the TV band.

From the standpoint of deriving objectives and evaluating performance, an especially troublesome problem centered around the inevitable small, random variations in the 60-cps ac power. This power line "bobble" creates "flicker" interference in TV signals which is most annoying at repetition rates around 5 cps.[7] Well-balanced video amplifier stages minimize sensitivity to additive components; cathode feedback and high-current operation of electron tubes minimize the nonlinear generation of bobble modulation components. However, to keep flicker interference within acceptable limits, the electronically regulated power supplies are counted on to reduce the effects of power line bobble by 50 db to 60 db. Electronic regulation plus dc operation of electron tube heaters and adequate filtering of unregulated power supplies control the amount of "hum" interference.

As a final result, the weighted interference to television by power line and other low-frequency noise in a terminal pair is expected to be 80 db below the peak-to-peak TV signal.

7.6 *Fluctuation Noise*

FM noise originating in the FM transmitter predominates over all other sources of fluctuation noise in FM terminals by at least 10 db. This noise originates in the two klystrons where it is generated by random (shot) processes in the electron beam.

Allowing for the frequency weighting of the interfering effect, for the action of typical TV circuit clampers and for the 12-db pre-emphasis, the klystrons introduce a fluctuation noise having the typical relative interference effect shown on Fig. 10(d). The integrated value of rms fluctuation noise in a terminal pair is 81 db below the peak-to-peak TV signal amplitude (8 mc peak-to-peak).

The relation of FM terminal noise to the over-all system noise for the telephone signal is discussed fully in Ref. 1. Briefly, noise interference

introduced by terminals is most serious in the telephone master-group at the lower end of the baseband. In a 3-kc band the rms frequency deviation due to noise is typically 35 cycles. Since an rms frequency deviation of 2.828 mc produces +8 dbm at the receiver output, the noise at this point is −90 dbm. At this point, the transmission level for the lowest level mastergroup is −25.5 db. Therefore, the noise is −64.5 dbm at 0 db TL; this is 17.5 dba in these "noisiest" telephone channels. If these channels remained at the same low-level end of the band (no frogging) for 16 terminals connected in tandem, the noise would add randomly to give a net noise meter reading of 29.5 dba at 0 db TL.

7.7 *Gain Stability*

Electronic stabilization of the most critical dc power supplies has reduced to insignificance the gain variations from these sources. However, power line voltage variation of ±1 per cent (typical stability) will cause ±0.2-db variation in the sensitivity of an FM receiver due to unregulated heater supplies. This random variation, added to systematic gain variations (estimated below as less than 0.05 db in one month), supports the expectation that gain changes of less than ±0.3 db will occur when an FM transmitter or an FM receiver is replaced by a standby unit through protection switching action in the field.

Presently available data on gain stability with time are meager, being the result of experimental observations over several month intervals on terminals in which the ages of all electron tubes were considerably under their life expectancy. Nevertheless, the performance under these condigave a reasonable degree of confidence in estimating that in a one month interval a typical terminal pair will have a systematic gain variation of −0.05 db and random variations not exceeding ±0.2 db. Sixteen terminal pairs in tandem would then have a net gain variation of −0.8 ± 0.8 db in a one-month interval.

APPENDIX A

Linearity, Differential Gain, Harmonic Distortion

Analytical expressions relating harmonic distortion to linearity and differential gain are obtained through the coefficients in the power series expansion for an output quantity (voltage, frequency deviation, etc.) as a function of input quantity. For example, consider the output vs. input characteristic

$$v_0 = a_1 v + a_2 v^2 + a_3 v^3 + \cdots, \qquad (1)$$

where the input, v, consists of a steady-state shift (or dc component), δ, added to a sinusoidal variation, $E \cos pt$; i.e., $v = \delta + E \cos pt$. Then, omitting dc terms, the output will contain the following sinusoidal components:

$$v_0 = F_1 \cos pt + F_2 \cos 2\, pt + F_3 \cos 3\, pt + \cdots$$

where

$$\begin{aligned} F_1 &= E[a_1 + 2a_2\delta + 3a_3\delta^2 + \cdots] \\ F_2 &= E^2 \left[\frac{a_2}{2} + \frac{3}{2}a_3\delta + \cdots\right] \\ F_3 &= E^3 \left[\frac{a_3}{4} + \cdots\right]. \end{aligned} \tag{2}$$

Differential gain is defined as the ratio of the small-signal gain (at fundamental frequency) for any shift, δ, to the gain when $\delta = 0$. Differential gain, as a function of δ, then has the form

$$\text{Differential gain} = 1 + L_1\delta + L_2\delta^2, \tag{3}$$

where, from the expression for F_1 in equation (2),

$$L_1 = \frac{2a_2}{a_1} \quad \text{and} \quad L_2 = \frac{3a_3}{a_1}.$$

Differential gain is normally expressed in db as 20 log $(1 + L_1\delta + L_2\delta^2)$.

Nonlinearity is closely related to differential gain and is defined as

$$\text{First-order nonlinearity} = L_1\delta \times 100 \text{ per cent} \tag{4}$$

$$\text{Second-order nonlinearity} = L_2\delta^2 \times 100 \text{ per cent.} \tag{5}$$

Harmonic performance can be expressed either in terms of the power series coefficients in (1) or the linearity coefficient in (3) as shown below:

$$\frac{\text{Second harmonic}}{\text{Fundamental}} = \frac{F_2}{F_1} = E\left[\frac{a_2}{2a_1} + \frac{3a_3}{2a_1}\delta - \left(\frac{a_2}{a_1}\right)^2 \delta \cdots\right]$$

The third term above is frequently negligible, in which case

$$\frac{F_2}{F_1} = E\left[\frac{1}{4} L_1 + \frac{1}{2} L_2\delta \cdots\right] \tag{6}$$

$$\begin{aligned} \frac{\text{Third harmonic}}{\text{Fundamental}} = \frac{F_3}{F_1} &= E^2\left[\frac{a_3}{4a_1} + \cdots\right] \\ &= E^2\left[\frac{1}{12} L_2 \cdots\right]. \end{aligned} \tag{7}$$

APPENDIX B

Klystron Quasi-Stationary Frequency Behavior

This appendix develops analytical relationships defining the FM properties of the 450A reflex klystron.

B.1 *Linearity of Steady-State Frequency Shift*

The detailed analysis of the behavior of reflex klystrons given in Ref. 10 suggests that a circuit analogue like that shown in Fig. 11 can be used to explain steady-state frequency behavior.

The phase shift, θ, in the drift space is a function of repeller voltage, v, and oscillating frequency, ω, while the phase shift, φ, in the resonator is a function of frequency only. To have sustained oscillations at any

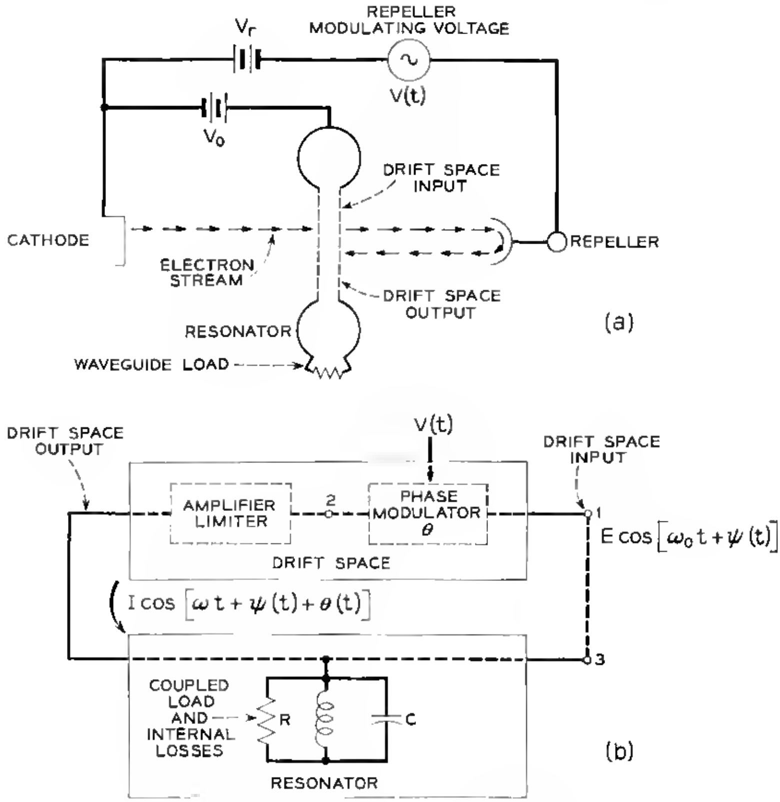

Fig. 11 — Circuit diagrams of reflex klystron: (a) schematic, (b) feedback loop.

frequency, the gain around the oscillator loop must be unity with a phase shift of zero. The equivalent amplifier-limiter action of the drift space takes care of satisfying the unity gain requirement. To satisfy the zero phase shift requirement, the oscillation frequency stabilizes at a value such that the resonator phase shift is equal and opposite to the drift space phase; i.e., $\varphi + \theta = 0$. If a change is made in the repeller voltage to $v + dv$ such that the delay in the drift space increases and the drift space phase changes to $\theta + d\theta$, then the oscillating frequency will decrease sufficiently to give an equal and opposite phase change in the resonator. Nonlinearity in the frequency-voltage relationship results primarily from the nonlinear relationship between frequency and phase in the resonator and to a lesser extent because the changes in drift space delay are not linearly related to changes in the repeller voltage. These relationships will be evident from solutions to the differential equation:

$$d\varphi + d\theta = 0 \tag{8}$$

from which

$$\frac{d\varphi}{d\omega} d\omega + \frac{\partial\theta}{\partial v} dv + \frac{\partial\theta}{\partial\omega} d\omega = 0 \tag{9}$$

and

$$\frac{d\omega}{dv} = -\frac{\dfrac{\partial\theta}{\partial v}}{\dfrac{d\varphi}{d\omega} + \dfrac{\partial\theta}{\partial\omega}}. \tag{10}$$

In Ref. 10, p. 593, the relationship between drift space phase and electrode voltages is given as

$$\theta = \frac{k\omega\sqrt{V_0}}{V_0 + V_r} \tag{11}$$

where the voltages V_0 and V_r are those shown in Fig. 11. Also, assuming that the resonator behaves as a simple parallel resonant circuit,

$$\varphi = \arc \tan Q \left[\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right]. \tag{12}$$

These relationships can be differentiated and the resulting expressions expanded in power series which are valid in the vicinity of the operating point of the klystron. When the first few terms of these series are substituted in (10), an expression for the small-signal FM sensitivity as a function of the frequency shift, $\delta = \omega - \omega_r$, and repeller voltage,

V_r, is obtained. The repeller voltage can be eliminated as a variable in the expression by substituting $V_r = V_{r0} + a\delta$, where $a = dv/d\omega$ evaluated at $\omega = \omega_r$ and V_{r0} is the repeller voltage corresponding to an oscillating frequency ω_r. The first few terms then become

$$\frac{d\omega}{dv} \approx \frac{\omega_r}{(V_0 + V_{r0})} \frac{1}{\left(\frac{2Q}{\theta_0} + 1\right)} \left[1 - \left(\frac{4Q}{\theta_0} - 1\right)\frac{\delta}{\omega_r} + \frac{8Q^3}{2Q + \theta_0}\frac{\delta^2}{\omega_r^2}\right], \quad (13)$$

where

$$\delta = \omega - \omega_r,$$

$$\omega_r = \text{resonant frequency of cavity in radians/sec,}$$

$$\theta_0 = \text{drift space phase angle in radians, at } \omega_r,$$

$$Q = \text{cavity } Q\text{, and}$$

$$V_0 + V_{r0} = \text{dc potentials applied to klystron as shown in Fig. 11.}$$

Comparing this expression with (3) we write

$$\text{Differential sensitivity} = 1 + L_1\delta + L_2\delta^2 \quad (14)$$

where

$$L_1 = -\left(\frac{4Q}{\theta_0} - 1\right) \Big/ \omega_r,$$

and

$$L_2 = \frac{8Q^3}{(2Q + \theta_0)\omega_r^2}.$$

Typical numerical values for these quantities are:

$$\omega_r = 2\pi \times 6 \times 10^9,$$

$$\theta_0 = 2\tfrac{3}{4} \text{ cycles} = 17.3 \text{ radians,}$$

$$Q = 100, \text{ and}$$

$$V_0 + V_r = 550;$$

which give:

$$\text{Differential sensitivity} = 1 - 22\frac{\delta}{\omega_r} + 37{,}000\frac{\delta^2}{\omega_r}.$$

For a $\pm$4-mc frequency deviation ($\delta/\omega_r = 4/6000$), the first-order non-

linearity is ± 1.5 per cent, and the second-order nonlinearity is 1.6 per cent. The first-order nonlinearity is primarily due to the nonlinear relationship between drift space phase shift and repeller voltage, while the second-order nonlinearity is predominently caused by the nonlinear phase shift in the resonator as a function of frequency.

B.2 *Second Harmonic Balance*

It is evident from the expression for F_2/F_1 in (6) that the second harmonic can be balanced to zero by selection of the operating point δ, such that $\delta = \delta_0 = -L_1/2L_2$. In terms of (14) a second harmonic balance can be obtained by means of a small change in the klystron repeller voltage so that the quiescent oscillating frequency is slightly different from the resonant frequency of the cavity, the amount of this difference being given by

$$\frac{\delta_0}{\omega_r} \approx \frac{\dfrac{4}{\theta_0} - 1}{\dfrac{16Q^3}{2Q + \theta_0}} = 3.0 \times 10^{-4}.$$

This corresponds to shifting to an operating point which is approximately 1.8 mc above the resonant frequency of the cavity.

B.3 *Experimental Adjustment for Second Harmonic Balance*

As can be checked by differentiating (13) and setting the resulting expression equal to zero, *minimum* differential sensitivity occurs at the same value of δ as was found for zero second harmonic. This principle is employed in a field adjustment procedure for "optimizing" klystron linearity: repeller bias is adjusted to minimize the small-signal FM deviation produced by a small voltage variation.

B.4 *Second-Order Balance Stability*

As a Function of Repeller Voltage. It has been shown above that it is possible to adjust the klystron bias so as to obtain a second harmonic balance. For such an adjustment to be worthwhile, however, the operating point must stay within bounds. For example, the ratio of the second harmonic to fundamental is given in (6) as

$$\frac{F_2}{F_1} = E\left(\frac{1}{4} L_1 + \frac{1}{2} L_2\delta\right),$$

where for $E = 4$ mc, the requirement on F_2/F_1 is -54 db* or 0.002. Substitution of the numerical values above gives a requirement on the frequency shift of 1.8 ± 1.0 mc. With a repeller sensitivity of about 1.45 mc per volt, the voltage stability requirement is ± 0.7 volt for the -100-volt repeller supply.

As a Function of the Resonator Voltage. The resonator has a sensitivity of approximately 0.5 mc per volt. Therefore, the frequency stability requirement of $+1.0$ mc, just derived, imposes a voltage requirement of about ± 2.0 volts on the 450-volt resonator supply. Thus, regulated supplies are required for both the repeller and resonator to maintain the required second-harmonic balance.

As a Function of Resonant Frequency. If temperature or tuning changes the resonant frequency of the cavity, the phase shift in the cavity is a function of two variables: the oscillating frequency, ω_0, and the resonant frequency, ω_r. With V constant, (9) becomes

$$\frac{\partial \varphi}{\partial \omega_r} d\omega_r + \frac{\partial \varphi}{\partial \omega_0} d\omega_0 + \frac{d\theta}{d\omega_0} d\omega_0 = 0$$

where

$$\frac{\partial \varphi}{\partial \omega_r} = -\frac{\partial \varphi}{\partial \omega_0}$$

for $\omega_0 \approx \omega_r$

Substituting, and solving for $d\omega_0/d\omega_r$ gives

$$\frac{d\omega_0}{d\omega_r} = \frac{\dfrac{\partial \varphi}{\partial \omega_0}}{\dfrac{\partial \varphi}{\partial \omega_0} + \dfrac{d\theta}{d\omega_0}}.$$

For the 450A klystron

$$\partial\varphi/\partial\omega_0 \approx \frac{2Q}{\omega_r} \approx 5.3 \times 10^{-9} \text{ second}$$

$$d\theta/d\omega_0 \approx \frac{\theta_0}{\omega_0} \approx 0.46 \times 10^{-9} \text{ second.}$$

Therefore,

$$\frac{d\omega_0}{d\omega_r} \approx 0.92$$

* Only a portion of the total requirement of —49 db is allocated to the klystron.

and as the resonant frequency changes, the oscillating frequency tends to follow so that the change in the desired 1.8-mc offset is only 8 per cent of the change in the resonant frequency.

Thermal and mechanical stabilities in the 450A klystron are such that second-order balance degradation due to rest frequency shifts caused by resonant frequency changes are generally negligible compared to the degradation due to resonator and repeller supply voltage changes.

B.5 *Third-Harmonic Distortion Performance*

The low-frequency third-harmonic performance of the 450A klystron with a resonator $Q = 100$ and for a peak frequency deviation of 4 mc is given by (7) with $E = 4$ mc and

$$L = \frac{8Q^3}{(2Q + \theta_0)\omega_r^2}.$$

Thus,

$$\frac{\text{Third harmonic}}{\text{Fundamental}} = E^2 \times \frac{8Q^3}{12\omega_r^2(2Q + \theta_0)}$$

$$= 0.00137 \text{ or } -57.2\ db.$$

This is 6 db better than the objective of -51 db.

APPENDIX C

Klystron Dynamic Behavior—Video Response

It was tacitly assumed in Appendix B that the steady-state frequency deviation behavior is also applicable when the incremental repeller voltage is a function of time. This assumption has been verified experimentally; frequency deviation is substantially independent of video signal frequency up to at least 10 mc. Similarly, harmonic distortion depends upon Q as predicted by the steady-state analysis.

This dynamic behavior may also be verified theoretically. The essence of this theoretical analysis will be outlined here to show how the integrating action of the resonator dynamically converts phase modulation to frequency modulation. Appendix D shows how this same action converts noise in the electron stream to an FM deviation noise in the output.

When a signal voltage is applied to the repeller in Fig. 11(b), a phase modulation of the carrier is produced in the closed loop. Finding a relationship between the phase modulation (output) and the modulating signal (input) is analogous to the problem encountered in any closed-loop

system such as a feedback amplifier. In principle, the loop between (1) and (3) on Fig. 11(b) is temporarily opened, and an arbitrarily amplitude and phase modulated signal,

$$e(t) = E_1(t) \cos [\omega_0 t + \psi(t)], \tag{15}$$

is introduced at 1. This signal is traced around the loop and appropriately modified until the output at 3 is expressed in terms of the arbitrary wave introduced at 1. At this point the input and output waves for the open loop can be equated to determine the closed-loop performance of the system. Since the output signal, as well as the input signal, will have the form of an amplitude and phase modulated wave, the amplitude modulation and phase modulation can be separately equated. Only the results for the phase modulation terms are of interest here. Some simplification is achieved by working in the frequency domain. Thus, the following definitions are made.

$S_\psi(\omega)$ = frequency spectrum of $\psi(t)$, and

$S_\theta(\omega)$ = frequency spectrum of $\theta(t)$, the phase modulation introduced by the modulating signal, $V(t)$.

The spectrum of the phase modulation at the output of the drift space becomes

$$S_\psi(\omega)e^{-j\omega D} + S_\theta(\omega)$$

where D equals the delay in drift space in seconds.

This spectrum is again modified as it passes through the resonator. The resonator has the effect of a low-pass filter on the phase modulation, where the filter is the low-pass equivalent of the actual bandpass structure. It thus consists of a resistor in parallel with a capacitor with values such that the bandwidth of the low-pass structure is just half that of the actual bandpass structure. For a Q of 100 at 6000 mc, the bandwidth is about 60 mc; the equivalent low-pass bandwidth is 30 mc, from which the low-pass transmission characteristic is therefore

$$Y(\omega) = \frac{1}{1 + j\omega\tau}, \tag{16}$$

with

$$\tau = 2Q/\omega_0 = 5.3 \times 10^{-9} \text{ second.}$$

The spectrum of the phase modulation at the output of the open loop is therefore

$$Y(\omega)S_\psi(\omega)e^{-j\omega D} + Y(\omega)S_\theta(\omega)$$

and under closed loop conditions

$$S_\psi(\omega) = Y(\omega)S_\psi(\omega)e^{-j\omega D} + Y(\omega)S_\theta(\omega) \tag{17}$$

from which

$$S_\psi(\omega) = \frac{Y(\omega)}{1 - Y(\omega)e^{-j\omega D}} S_\theta(\omega). \tag{18}$$

Substitution for $Y(\omega)$ as given in equation (16) permits the excellent approximation,

$$S_\psi(\omega) \approx \frac{1}{j\omega(\tau + D)} S_\theta(\omega) \tag{19}$$

which for the 450A klystron is accurate to within about 0.01 db up to the top baseband frequency of 10 mc. After both sides of (19) are multiplied by $j\omega$, the following identifications are made:

$j\omega S_\psi(\omega)$ = frequency spectrum of $\psi'(t)$, the first time derivative of $\psi(t)$, and

$S_\theta(\omega)$ = frequency spectrum of $\theta(t)$.

Since the instantaneous frequency of a wave is defined as the first time derivative of the instantaneous phase, the time domain equivalent of (19) can be written as

$$\text{frequency modulation} = \psi'(t) = \frac{\theta(t)}{\tau + D}. \tag{20}$$

Thus, within the accuracy of the approximation stated above, the closed-loop frequency modulation is directly proportional to the phase modulation introduced by the modulating signal. Furthermore, the equivalence of this result to the quasi-stationary result obtained in Appendix B is possible when the following identifications are made:

$$\tau = \frac{2Q}{\omega_r} = \frac{d\varphi}{d\omega}$$

$$D = \frac{\partial\theta}{\partial\omega}.$$

Thus, the above result can also be written in the form

$$\varphi'(t) = \frac{\theta(t)}{\dfrac{d\varphi}{d\omega} + \dfrac{\partial\theta}{\partial\omega}} \approx \frac{\dfrac{\partial\theta}{\partial v}}{\dfrac{d\varphi}{d\omega} + \dfrac{\partial\theta}{\partial\omega}} v(t)$$

which is the linearized dynamic equivalent of (10).

From the preceding analysis the following conclusions are obtained:

(a) Phase modulation is dynamically converted to frequency modulation by regenerative action in the feedback loop.

(b) Klystron modulation sensitivity, inversely proportional to open loop delay, is the same as that found in the steady-state analysis.

(c) The resultant frequency deviation is essentially independent of video signal frequency.

APPENDIX D

Klystron Fluctuation noise

A comparison of the FM noise characteristic of 450A klystrons, Fig. 10(d), with noise characteristics typical of other electron tube devices leads to the conclusion that the two phenomena are identifiable. "Flicker" noise, varying approximately as $1/f$, predominates for frequencies less than 25 kc. "Shot" noise, flat with frequency, predominates at frequencies greater than 25 kc. Since shot noise is controlling over all of the message band and a major portion of the television band, it is the more important source of klystron noise.

Flicker noise is associated with random time variations in group electron emission from the cathode. The power spectrum for these variations decreases at about a $1/f$ rate. The exact mechanism whereby this random amplitude modulation of the dc space current is converted to frequency modulation of the carrier has not been quantitatively identified. It probably comes from variations in the inter-action grid capacitance or drift space delay induced by the space charge density modulation.

On the other hand, shot noise has been clearly identified as being due to the statistical time variations for single electron emissionor (capture by intervening grids), which have spectral intensities that are relatively constant up to and above 6000 mc. The electron stream flowing through the resonator interaction grids [twice, as illustrated in Fig. 11(a)] varies randomly with time. Those frequency components of the variations which fall in bands equally displaced on either side of the carrier add to it, and randomly vary its phase. This phase variation is converted to random frequency deviation by the action of the regenerative loop as described in the previous section.

Shot noise per cycle of bandwidth is given by the following equation for temperature-limited emission:

$$i_s^2 = 2(3.18 \times 10^{-19})\, I \tag{21}$$

where i_s is the rms fluctuation current and I the dc beam current. The

factor 2 is used to allow for the double transit of the electron stream through the interaction grids.

Also flowing into the interaction grid space is a bunched electron stream which has an equivalent fundamental carrier current component, i_c, at radian frequency, ω_0. This current produces power, P, in the resonator load. Therefore,

$$i_c = \sqrt{\frac{P}{R}} = \sqrt{\frac{P\omega_0 C}{Q}} \tag{22}$$

where

R = resistance of the loaded resonator, and

C = capacitance of the interaction grid space.

This carrier current will be randomly phase modulated by upper and lower sideband noise currents to give a net open-loop rms phase deviation in each one cps band of

$$\theta_n = \frac{i_s}{i_c} = i_s \sqrt{\frac{Q}{P\omega_0 C}} \text{ radians}. \tag{23}$$

Regenerative action described in Appendix C will change this to a frequency deviation which is independent of the noise spectrum frequency. Neglecting D with respect to τ in equation (20):

$$\begin{aligned} \text{rms frequency deviation} &= \frac{\theta_n}{\tau} = \frac{\theta_n \omega_0}{2Q} \\ &= i_s \sqrt{\frac{\omega_0}{4PQC}} \text{ rad/sec in a one-cps band due to noise.} \end{aligned} \tag{24}$$

Values typical of the 450A klystron are:

$2I$ = 0.1 amp,

P = 0.2 watt,

C = 0.5 micromicrofarad,

Q = 100, and

$\omega_0 = 2\pi \times 6175 \times 10^6$.

When substituted in (24), these give an rms frequency deviation in a one-cps band due to shot noise of 0.9 cps.

Experimentally, it is found that one 450A klystron generates an rms

frequency deviation of about 0.4 cps in a one-cps band*; for the two klystrons in the FM modulator the noise is 1.41 times greater.

Apparently space-charge smoothing, which usually reduces rms shot noise in space-charge-limited devices by a factor of 5, gives a 2/1 improvement in this klystron. Alternatively, space-charge smoothing may be more fully effective, and partition capture may be responsible for the added noise.

APPENDIX E

Discriminator Network Design

To demonstrate analytically the main features of the discriminator network design, the magnitudes of the interstage impedances for the common interstage and for the high- and low-frequency discriminator networks as shown in Fig. 7 are expanded in power series about the center frequency of the discriminator. For the common interstage this gives

$$| Z_c | = | Z_{c0} | (1 + c_1\delta + c_2\delta^2 + c_3\delta^3 \cdots), \tag{25}$$

where

$$\delta = \omega - \omega_0 \text{, and}$$

$$\omega_0 = \text{center frequency of the discriminator.}$$

Similarly, for the high- and low-frequency discriminator networks,

$$| Z_l | = | Z_{l0} | (1 + l_1\delta + l_2\delta^2 + l_3\delta^3 + \cdots) \tag{26}$$

$$| Z_h | = | Z_{h0} | (1 + h_1\delta + h_2\delta^2 + h_3\delta^3 + \cdots). \tag{27}$$

The amplitude characteristics for transmission through the two sides of the discriminator are then obtained as the product of $| Z_c |$ with $| Z_l |$ and $| Z_h |$ respectively. Thus,

$$A_l = k_l | Z_c | | Z_l | \tag{28}$$

$$A_h = k_h | Z_c | | Z_h | \tag{29}$$

where the factors k_l and k_h are introduced to include the effect of electron tube gains in the two sides.

For the purpose of the following discussion the constant, c_1, will be taken equal to zero. This is done in the actual design by tuning the in-

* To generate this same noise by thermal agitation would require a resistance in the repeller circuit of 5 megohms; the actual circuit value is 800 ohms.

terstage network to the center frequency of the discriminator. Performing the multiplications indicated in (28) and (29) and discarding terms higher than δ^3 gives

$$A_l = A_{l0}[1 + l_1\delta + (l_2 + c_2)\delta^2 + (l_3 + c_3 + l_1c_2)\delta^3] \tag{30}$$

$$A_h = A_{h0}[1 + h_1\delta + (h_2 + c_2)\delta^2 + (h_3 + c_3 + h_1c_2)\delta^3] \tag{31}$$

where

$$A_{l0} = k_l \mid Z_{c0} \mid \mid Z_{l0} \mid \tag{32}$$

$$A_{h0} = k_h \mid Z_{c0} \mid \mid Z_{h0} \mid . \tag{33}$$

The input signal has the form,

$$i(t) = i_0[1 + A(t)] \cos [\omega_0 t + \varphi(t)] \tag{34}$$

where

$$A(t) = \text{amplitude modulation,}$$

$$\varphi(t) = \text{phase modulation, and}$$

$$\varphi'(t) = \text{frequency modulation.}$$

As a result of the amplitude shape in the discriminator networks, the instantaneous amplitudes at the output of the networks (or input to AM detectors) are approximately* as follows:

$$M_l(t) = i_0 A_{l0}[1 + A(t)]$$
$$[1 + l_1\varphi' + (l_2 + c_2)\varphi'^2 + (l_3 + c_3 + l_1c_2)\varphi'^3] \tag{35}$$

$$M_h(t) = i_0 A_{h0}[1 + A(t)]$$
$$[1 + h_1\varphi' + (h_2 + c_2)\varphi'^2 + (h_3 + c_3 + h_1c_2)\varphi'^3] \tag{36}$$

The output of the balanced discriminator is given by the difference between the outputs of the two detectors. If the diode detector efficiencies for the two sides of the discriminator are D_l and D_h, the discriminator output is given as

$$\text{Discriminator output} = D_h M_h(t) - D_l M_l(t). \tag{37}$$

* These results are based on a quasi-stationary approach and are valid for very low modulating frequencies. At higher modulating frequencies the relationship between instantaneous frequency and instantaneous amplitude is more complex. The additional terms which then arise are due primarily to delay distortion which is compensated by delay equalization (see Section 7.2).

The individual output terms are listed and discussed below. For simplicity in the expressions which have to be written we let $E_l = i_0 A_{l0} D_l$ and $E_h = i_0 A_{h0} D_h$.

(i) $E_h - E_l =$ dc output.

(ii) $(E_h - E_l)A(t) =$ output due to unsuppressed amplitude modulation. By adjusting the relative gains in two sides of the discriminator so that the dc output is zero, this term is also eliminated.

(iii) $(E_h h_1 - E_l l_1)\varphi'(t) =$ desired signal output. Since h_1 is positive and l_1 is negative, the signal components from the two sides actually add in the output.

(iv) $[E_h(h_2 + c_2) - E_l(l_2 + c_2)]\varphi'(t)^2 =$ unwanted second-order modulation. This term will be discussed in detail later.

(v) $[E_h(h_3 + c_3 + h_1 c_2) - E_l(l_3 + c_3 + l_1 c_2)]\varphi'(t)^3 =$ unwanted third-order modulation.

(vi) Finally there is a set of terms identical to (iii), (iv), and (v) above except that each is multiplied by $A(t)$. By keeping $A(t)$ small by means of limiter action ahead of the discriminator, these terms are held to acceptable levels.

The effect of the common interstage is demonstrated by examination of the distortion term listed in (iv) above. In the absence of the common interstage, this term would be

$$[E_h h_2 - E_l l_2]\varphi'(t)^2$$

and, with $E_h = E_l$ to minimize distortion term (ii), it is desirable, and possible, to adjust the discriminator so that $h_2 = l_2$ and the term goes to zero. For the values of h_2 and l_2 of the actual discriminator design, the requirement on the stability of E_h and E_l to meet the modulation objective would then be given by

$$| E_h - E_l | \leqq \frac{| E_h |}{15}.$$

This corresponds to holding the relative gains of the two sides of the discriminator equal to within about 0.5 db between maintenance intervals.

The use of a common interstage selected so that $c_2 = -h_2 = -l_2$ causes the distortion term to go to zero even though E_h and E_l are not equal. This is illustrated in Fig. 9(a). In practice, it has been found

possible to adjust c_2 such that $c_2 + h_2 = c_2 + l_2$ are about 10 per cent of $h_2 = l_2$, which relaxes the requirement on gain stability to

$$| E_h - E_l | \leqq \frac{| E_h |}{1.5},$$

which corresponds to a relative gain stability of 4.4 db.

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